METHOD AND ARRANGEMENT FOR NOISE VARIANCE AND SIR ESTIMATION

5 **Field of the Invention**

This invention relates to noise variance and signal/interference ratio (SIR) estimation, and particularly though not exclusively to such estimation in

10 wireless communication receivers. It will be understood that, as used herein, the terms 'noise' and 'interference' are to be considered synonymous, with each encompassing both noise and interference.

15

Background of the Invention

In the field of this invention it is known that many parts of a wireless communications receiver often require

20 an estimation of noise variance and/or SIR. This is needed for purposes of power control, threshold determination for various algorithms, quantisation of soft-decision information for channel decoding purposes to name but a few.

25

For BPSK (Binary Phase Shift Key) and QPSK (Quadrature Phase Shift Key) modulation the conventional method for estimating the SIR at the output of a detector relies on estimating output noise variance using the following

30 equality known (for example) from the publication by

Papoulis and Pillai, entitled 'Probability, Random
Variables and Stochastic Processes', 3rd Ed. 1991,

$$\hat{\sigma}_z^2 = E\left(\left|\hat{d}_n^{(k)}\right|^2\right) - E\left(\left|\hat{d}_n^{(k)}\right|\right)^2$$

where $\hat{\sigma}_z^2$ represents variance, E represents mean value

and $\hat{d}_n^{(k)}$ represents the detector output.

This yields the following result:

$$SIR^{(k)} = \frac{P^{(k)} - \hat{\sigma}_z^2}{\hat{\sigma}_z^2} = \frac{E\left(\left|\hat{d}_n^{(k)}\right|\right)^2}{E\left(\left|\hat{d}_n^{(k)}\right|^2\right) - E\left(\left|\hat{d}_n^{(k)}\right|\right)^2}$$

where SIR represents the SIR of the kth sequence at the

detector output, and $P^{(k)}$ represents the average power of

the kth sequence at the detector output.

However, this approach has the disadvantage(s) that the
accuracy of this method at low SIR is poor since it
suffers from a bias term. An analysis of the bias term
and a correction method has been suggested in GB patent
application no. 0128475.1. However, the suggested
correction method requires a look-up table to correct for
the aforementioned problem, and the estimation variance
is also increased when correcting the bias.

A need therefore exists for a method and arrangement for
noise variance and SIR estimation wherein the
abovementioned disadvantage(s) may be alleviated.

Statement of Invention

In accordance with a first aspect of the present
invention there is provided a method for noise variance
5   estimation as claimed in claim 1.

In accordance with a second aspect of the present
invention there is provided an arrangement for noise
variance estimation as claimed in claim 10.

10

Preferably, the second noise variance signal is produced
by applying to the first noise variance signal a function
substantially equal to the detector's transfer function.

15   Preferably, the first noise variance signal is derived
from a midamble portion of the received signal.

Preferably, an estimate of total power at the detector
output is produced from the second noise variance signal
20   and an SIR signal representative of SIR in the received
signal.

Brief Description of the Drawings

25

One method and arrangement for noise variance and SIR
estimation incorporating the present invention will now
be described, by way of example only, with reference to
the accompanying drawings, in which:

30

FIG. 1 shows a block schematic diagram illustrating a 3GPP radio communication system in which the present invention may be used;

5   FIG. 2 shows a block schematic diagram illustrating a generic burst structure employed in the system of FIG. 1; and

FIG. 3 shows a block schematic diagram illustrating
10   a detector arrangement incorporating a noise variance and SIR estimator utilising the present invention.

15 **Description of Preferred Embodiment(s)**

The following preferred embodiment of the present invention will be described in the context of a 3GPP (3rd Generation Partnership Project) UMTS (Universal Mobile
20 Telecommunication System) Radio Access Network (UTRAN) system operating in TDD mode. Referring firstly to FIG. 1, a typical, standard UTRAN system 100 is conveniently considered as comprising: a terminal/user equipment domain 110; a UMTS Terrestrial Radio Access
25 Network domain 120; and a Core Network domain 130.

In the terminal/user equipment domain 110, terminal equipment (TE) 112 is connected to mobile equipment (ME) 114 via the wired or wireless R interface. The ME 114 is
30 also connected to a user service identity module (USIM) 116; the ME 114 and the USIM 116 together are considered

as a user equipment (UE) 118. The UE 118 communicates
data with a Node B (base station) 122 in the radio access
network domain 120 via the wireless *Uu* interface. Within
the radio access network domain 120, the Node B 122

5   communicates with a radio network controller (RNC) 124
via the *Iub* interface. The RNC 124 communicates with other
RNC's (not shown) via the *Iur* interface. The Node B 122
and the RNC 124 together form the UTRAN 126. The RNC 124
communicates with a serving GPRS service node (SGSN) 132

10   in the core network domain 130 via the *Iu* interface.
Within the core network domain 130, the SGSN 132
communicates with a gateway GPRS support node (GGSN) 134
via the *Gn* interface; the SGSN 132 and the GGSN 134
communicate with a home location register (HLR) server

15   136 via the *Gr* interface and the *Gc* interface
respectively. The GGSN 134 communicates with public data
network 138 via the *Gi* interface.

Thus, the elements RNC 124, SGSN 132 and GGSN 134 are

20   conventionally provided as discrete and separate units
(on their own respective software/hardware platforms)
divided across the radio access network domain 120 and
the core network domain 130, as shown the FIG. 2.

25   The RNC 124 is the UTRAN element responsible for the
control and allocation of resources for numerous Node B's
122; typically 50 to 100 Node B's may be controlled by
one RNC. The RNC also provides reliable delivery of user
traffic over the air interfaces. RNC's communicate with

30   each other via the *Iur* interface.

The SGSN 132 is the UMTS Core Network element responsible
for Session Control and interface to the HLR. The SGSN
keeps track of the location of an individual UE and
performs security functions and access control. The SGSN
5   is a large centralised controller for many RNCs.

The GGSN 134 is the UMTS Core Network element responsible
for concentrating and tunnelling user data within the
core packet network to the ultimate destination (e.g.,
10 internet service provider - ISP).

Such a UTRAN system and its operation are described more
fully in the 3GPP technical specification documents
3GPP TS 25.401, 3GPP TS 23.060, and related documents,
15 available from the 3GPP website at www.3gpp.org, and need
not be described in more detail herein.

The physical layer of UTRA TDD mode provides physical
channels that carry transport channels from the MAC
20 (Medium Access Control) sub-layer of UMTS Layer 2. A
physical channel is defined by frequency, timeslot,
channelisation code, burst type, and radio frame
allocation. In UMTS Layer 2, in each time slot, three
burst structures (as shown generically in FIG. 2) can be
25 supported, each consisting of two data fields (210, 230),
a midamble (220) and a guard period (240).

The data fields contain the data symbols from the
transport channels, after the processes of coding,
30 multiplexing, interleaving, and modulation. The midamble
field contains the training sequence, which is used in a

number of Layer 1 algorithms, such as channel estimation. The guard period, GP, is used to accommodate any timing inaccuracies, from propagation delays, channel dispersion, and power ramping in the transmitter. The different burst types and their associated field lengths in chips are given in the table below:

Burst Type	Data Field 1	Data Field 2	Midamble L_m	GP
Burst Type 1	976	976	512	96
Burst Type 2	1104	1104	256	96
Burst Type 3	976	880	512	192

The received sequence in the data payload areas of the burst is given by

$$\mathbf{e} = \mathbf{A}\mathbf{d} + \mathbf{n}$$

where

$$\mathbf{d} = \left(\mathbf{d}^{(1)T}, \mathbf{d}^{(2)T}, \ldots, \mathbf{d}^{(K)T}\right)^T = \left(d_1, d_2, \ldots, d_{KN}\right)^T,$$

$(.)^T$ denotes transposition, K is the number of data sequences $k = 1, \ldots, K$, and N is the number of symbols per data sequence $n = 1, \ldots, N$. The data sequence for the kth user is given by $\mathbf{d}^{(k)} = \left(d_1^{(k)}, d_2^{(k)}, \ldots, d_N^{(k)}\right)^T$. The noise sequence $\mathbf{n}$, denoted by $\mathbf{n} = \left(n_1, n_2, \ldots, n_{NQ+W-1}\right)^T$, has zero mean and covariance matrix $\mathbf{R}_n = E\left(\mathbf{n}\mathbf{n}^H\right)$, where $(.)^H$ denotes conjugate transposition. The matrix has dimensions $(NQ + W - 1) \times KN$ and the elements are given by:

$$A_{Q(n-1)+l, N(k-1)+n} = \begin{cases} \mathbf{b}^{(k)} & k = 1 \ldots K, \ l = 1 \ldots Q + W - 1, \ n = 1 \ldots N \\ 0 & else \end{cases}$$

where $\mathbf{b}^{(k)} = \mathbf{h}^{(k)} * \mathbf{c}^{(k)}$, $\mathbf{h}^{(k)}$ is the impulse response of the kth user, W is the length of the channel impulse response, $*$ denotes discrete time convolution, $\mathbf{c}^{(k)}$ is the spreading code of the kth user and Q is the length of the spreading sequence in chips.

The output of the detector is given by

$$\hat{\mathbf{d}} = f(\mathbf{e}) = \mathbf{r} + \mathbf{z}$$

where $f(.)$ denotes the transfer function of the detector, the vector $\mathbf{r}$ contains the desired symbols, and the vector $\mathbf{z}$ contains noise plus interference. The average power for the kth sequence at the output of the detector is given by

$$P^{(k)} = E\left(\left|\hat{d}_n^{(k)}\right|^2\right)$$

Expanding $P^{(k)}$ produces

$$P^{(k)} = E\left(\left|r_{n+(k-1)Q}\right|^2\right) + E\left(r_{n+(k-1)Q}z^*_{n+(k-1)Q}\right) + E\left(r^*_{n+(k-1)Q}z_{n+(k-1)Q}\right) + E\left(\left|z_{n+(k-1)Q}\right|^2\right)$$

Under the assumption that the noise is uncorrelated with the signal vector $\mathbf{r}$, the average power for the kth sequence becomes

$$P^{(k)} = E\left(\left|r_{n+(k-1)Q}\right|^2\right) + \sigma_z^2$$

where $E(.)$ is the statistical average, $\sigma_z^2 = E\left(\left|z_{n+(k-1)Q}\right|^2\right)$ is the noise variance at the output of the detector, and $E\left(\left|r_{n+(k-1)Q}\right|^2\right)$ is the signal power for the kth sequence. The SIR at the output of the detector for the kth sequence is therefore given by

$$SIR^{(k)} = \frac{P^{(k)} - \sigma_z^2}{\sigma_z^2} = \frac{E\left(\left|r_{n+(k-1)\varrho}\right|\right)}{\sigma_z^2}$$

The conventional method for estimating the SIR relies on estimating the detector output noise variance using the following equality mentioned above:

$$\hat{\sigma}_z^2 = E\left(\left|\hat{d}_n^{(k)}\right|^2\right) - E\left(\left|\hat{d}_n^{(k)}\right|\right)^2$$

to yield the following result:

$$SIR^{(k)} = \frac{P^{(k)} - \hat{\sigma}_z^2}{\hat{\sigma}_z^2} = \frac{E\left(\left|\hat{d}_n^{(k)}\right|\right)^2}{E\left(\left|\hat{d}_n^{(k)}\right|^2\right) - E\left(\left|\hat{d}_n^{(k)}\right|\right)^2}$$

10   As discussed above, the accuracy of this approach at low
     SIR is poor since it suffers from a bias term, which may
     be corrected by use of a look-up table.

     As will be discussed in greater detail below, the
15   following preferred embodiments of the present invention
     do not suffer from such a bias term and therefore do not
     require a look-up table to correct for the aforementioned
     problem.

20   Referring now to FIG. 3, in detector arrangement 300
     (which may be used in Node B 122 or User Equipment 118) a
     new technique for estimating the SIR at the output of a
     detector 310 uses an estimate of the detector input noise
     variance (derived in channel estimator 320) to provide an
25   estimate of the detector output noise variance (derived
     in SIR estimator 330). In the channel estimator 320 the
     detector input noise variance is derived from the

midamble portion of the burst using locally reconstructed
replicas of the transmitted signal which are removed from
the received set of samples providing a residual term
that contains the noise of the channel. In the SIR

5   estimator 330, by deriving the transfer function of the
detector an estimate of the detector output noise
variance is estimated.

The estimated output noise variance then allows an

10   improved estimate of the SIR ($SIR^{(1)}...SIR^{(K)}$) at the detected
output. Typically, the SIR at the output of the detector
is used for deriving soft decision quantisation levels
for application to channel decoding algorithms.

15   In the following description, two types of CDMA (Code
Division Multiple Access) detector are considered, namely
single user detector (SUD) and multiuser detector (MUD).
It will be understood that the invention is also
applicable to other types of detector such as a RAKE

20   receiver.

The technique described here is based on first estimating
the noise variance at the input to the detector and then
mapping the input noise variance to the output noise

25   variance using the transfer function of the detector.

The process of estimating the noise variance at the input
to the detector is carried out using the midamble portion
of the burst. Considering the received sequence of chip

30   spaced samples $\mathbf{e} = (e_1, e_2, ..., e_{L_B})^T$, where L_B is the burst

length, a locally reconstructed version for the midamble
portion of the burst is defined as follows:

$$\overline{\mathbf{e}} = \left(\overline{e}_1, \overline{e}_2, \ldots, \overline{e}_{L_m + W - 1}\right)^T = \left(\sum_k \left(\mathbf{h}^{(k)} * \mathbf{m}^{(k)}\right)\right)^T$$

where $\overline{}$ denotes locally reconstructed version and $\mathbf{m}^{(k)}$ is

the midamble sequence of the kth user. The noise variance
estimate at the input to the detector is given by

$$\hat{\sigma}^2 = E\left(\left\|\left(e_W, e_{W+1}, \ldots, e_X\right) - \left(\overline{e}_W, \overline{e}_{W+1}, \ldots, \overline{e}_X\right)\right\|^2\right)$$

where $X \le L_m$ and the starting position is W since the

first $W - 1$ samples from the midamble portion of the burst

are corrupted by the data portion of the burst.

Multiuser Detection

Under the assumption that the noise is white with

variance σ^2, the Minimum Mean Squared Error (MMSE) block
linear equalizer solution to symbol estimation is given
(as known from the publication of Klein, Kaleh and Baier
entitled 'Zero Forcing and Minimum Mean-Square-Error
Equalization for Mutliuser Detection in Code-Division
Multiple-Access Channels' in IEEE Trans VT, VOL.45, No.2,
May 1996, pp276-287) by

$$\hat{\mathbf{d}} = \left(\mathbf{A}^H \mathbf{A} + \sigma^2 \mathbf{I}\right)^{-1} \mathbf{A}^H \mathbf{e} = f(\mathbf{e}) = \mathbf{r} + \mathbf{z}$$

where $\mathbf{I}$ is the identity matrix and

$$\mathbf{r} = \left(\mathbf{A}^H \mathbf{A} + \sigma^2 \mathbf{I}\right)^{-1} \mathbf{A}^H \mathbf{A} \mathbf{d} = \left(r_1, r_2, \ldots r_{KN}\right)^T \qquad (1)$$

$$\mathbf{z} = \left(\mathbf{A}^H \mathbf{A} + \sigma^2 \mathbf{I}\right)^{-1} \mathbf{A}^H \mathbf{n} = \left(z_1, z_2, \ldots, z_{KN}\right)^T \qquad (2)$$

From equation (2), the noise variance seen at the output of the MUD is given by

$$\sigma_z^2 = E\left(\left|z_{n+(k-1)Q}\right|^2\right) = \left(\left\|\mathbf{b}^{(k)}\right\|^2 + \sigma^2\right)^{-2} \left\|\mathbf{b}^{(k)}\right\|^2 \sigma^2$$

where $\|\|$ denotes vector norm, and σ^2 represents the noise

5   variance at the input of the MUD. By replacing σ^2 with the estimate of the MUD input noise variance $\hat{\sigma}^2$, we have a direct method for estimating the MUD output noise variance $\hat{\sigma}_z^2$. For completeness, the estimate $\hat{\sigma}_z^2$ of the MUD output noise variance can be written as

$$\hat{\sigma}_z^2 = f\left(\hat{\sigma}^2\right) = \left(\left\|\mathbf{b}^{(k)}\right\|^2 + \hat{\sigma}^2\right)^{-2} \left\|\mathbf{b}^{(k)}\right\|^2 \hat{\sigma}^2$$

10

where $f\left(\hat{\sigma}^2\right)$ represents the noise transfer function of the detector.

Using the new estimate for the output noise variance, the

15   SIR at the output of the MUD for the kth sequence is defined by

$$SIR^{(k)} = \frac{E\left(\left|r_{n+(k-1)Q}\right|^2\right) + \delta\left(\hat{\sigma}^2\right)}{\hat{\sigma}_z^2}$$

where the error term $\delta\left(\hat{\sigma}^2\right)$ is given by

$$\delta\left(\hat{\sigma}^2\right) = \left\|\mathbf{b}^{(k)}\right\|^2 \left(\frac{\sigma^2}{\left(\left\|\mathbf{b}^{(k)}\right\|^2 + \sigma^2\right)^2} - \frac{\hat{\sigma}^2}{\left(\left\|\mathbf{b}^{(k)}\right\|^2 + \hat{\sigma}^2\right)^2} \right)$$

20

It is clear from the above set of equations that when $\hat{\sigma}^2 = \sigma^2$ we have the following

$$SIR^{(k)} = \frac{E\left(\left|r_{n+(k-1)Q}\right|^2\right)}{\sigma_z^2}$$

It will therefore be understood that the accuracy of the above technique is directly related to the quality of the

5 noise variance estimate, $\hat{\sigma}^2$, at the input of the MUD.

Single User Detection

For the single user detector case the received sequence

10 is written as

$$\mathbf{e} = \mathbf{Ad} + \mathbf{n} = \mathbf{HCd} + \mathbf{n}$$

The matrix **H** has dimensions $(NQ+W-1)\times NQ$ and its elements are given by

15
$$H_{i,v} = \begin{cases} h_{i-v+1} & 1 \le i-v+1 \le W \\ 0 & else \end{cases}$$

where $\mathbf{h} = (h_1, h_2, \ldots, h_W)^T$, $i = 1, \ldots, NQ+W-1$, and $v = 1, \ldots, NQ$. The

matrix **C** has dimensions $NQ \times KN$ and its elements are given by

$$C_{Q(n-1)+q, N(k-1)+n} = \begin{cases} c_q^{(k)} & q = 1, \ldots, Q, n = 1, \ldots, N, k = 1, \ldots, K \\ 0 & else \end{cases}$$

20

For Minimum Mean Squared Error (MMSE) symbol estimation and under the assumption that the noise is white with variance σ^2, the output of the SUD is given by

$$\hat{\mathbf{d}} = \mathbf{C}^H \left(\mathbf{H}^H \mathbf{H} + \sigma^2 \mathbf{I}\right)^{-1} \mathbf{H}^H \mathbf{e} = f(\mathbf{e}) = \mathbf{r} + \mathbf{z}$$

25 where

$$\mathbf{r} = \mathbf{C}^H \left(\mathbf{H}^H \mathbf{H} + \sigma^2 \mathbf{I}\right)^{-1} \mathbf{H}^H \mathbf{HCd} = (r_1, r_2, \ldots r_{KN})^T \tag{3}$$

$$z = C^H \left(H^H H + \sigma^2 I\right)^{-1} H^H n = \left(z_1, z_2, \ldots, z_{KN}\right)^T \qquad (4)$$

From equation (4), the noise variance seen at the output of the SUD is given by

$$\sigma_z^2 = E\left(\left|z_{n+(k-1)Q}\right|^2\right) = G \times \left(\|h\|^2 + \sigma^2\right)^{-2} \|h\|^2 \sigma^2$$

where $\|\|$ denotes vector norm, the multiplier G comes from the matrix C, and in general $G = \left\|c^{(k)}\right\|^2 = Q$.

By replacing σ^2 with the estimate of the SUD input noise variance $\hat{\sigma}^2$, we have a direct method for estimating the SUD output noise variance σ_z^2. For completeness, the estimate $\hat{\sigma}_z^2$ of the SUD output noise variance can be written as

$$\hat{\sigma}_z^2 = f\left(\hat{\sigma}^2\right) = Q \times \left(\|h\|^2 + \hat{\sigma}^2\right)^{-2} \|h\|^2 \hat{\sigma}^2$$

where G is replaced with Q and $f\left(\hat{\sigma}^2\right)$ is the noise transfer function of the detector. Using the new estimate for the output noise variance, the SIR at the output of the SUD for the kth sequence is defined by

$$SIR^{(k)} = \frac{E\left(\left|r_{n+(k-1)Q}\right|^2\right) + \delta\left(\hat{\sigma}^2\right)}{\hat{\sigma}_z^2}$$

where the error term $\delta\left(\hat{\sigma}^2\right)$ is given by

$$\delta\left(\hat{\sigma}^2\right) = Q \times \|h\|^2 \left(\frac{\sigma^2}{\left(\|h\|^2 + \sigma^2\right)^2} - \frac{\hat{\sigma}^2}{\left(\|h\|^2 + \hat{\sigma}^2\right)^2}\right)$$

It is clear from the above set of equations that when $\hat{\sigma}^2 = \sigma^2$ we have the following

$$SIR^{(k)} = \frac{E\left(\left|r_{n+(k-1)Q}\right|^2\right)}{\sigma_z^2}$$

It will therefore be understood that the accuracy of the
above technique is directly related to the quality of the

5 noise variance estimate, $\hat{\sigma}^2$, at the input of the SUD.

It will be appreciated that the method described above
for noise variance and SNIR estimation may be carried out
in software running on a processor (not shown - e.g., in

10 User Equipment such as 118 or a Node B such as 122), and
that the software may be provided as a computer program
element carried on any suitable data carrier (also not
shown) such as a magnetic or optical computer disc.

15 It will be also be appreciated that the arrangement
described above for noise variance and SNIR estimation
may alternatively be carried out in hardware, for example
in the form of an integrated circuit (not shown) such as
an FPGA (Field Programmable Gate Array) or ASIC

20 (Application Specific Integrated Integrated Circuit).

It will be understood that the method and arrangement for
noise variance and SIR estimation described above
provides the following advantages that the accuracy of

25 this technique is not poor at low SIR, since it does not
suffer from a bias term, nor does it require correction
therefor using a look-up table. An additional advantage
is that any increase in estimation variance resulting
from bias correction may be avoided.

30